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Dattatreya

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(54) **RECEPTION OF PHASE-SHIFTED SIGNAL SEQUENCES IN DIVERSITY RECEPTION**

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H04L 27/26 (2006.01)

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CPC ***H04L 27/2275*** (2013.01); ***H04L 27/2613*** (2013.01)

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See application file for complete search history.

(56) **References Cited**

U.S. PATENT DOCUMENTS

5,311,523	A	5/1994	Serizawa et al.	
7,050,001	B2	5/2006	Buchler	
8,059,698	B2 *	11/2011	Mester et al.	375/150
8,301,037	B2	10/2012	Pfau	
8,654,911	B2 *	2/2014	Van Nee	375/347
8,675,754	B1 *	3/2014	Yonge et al.	375/261
2003/0012267	A1 *	1/2003	Jitsukawa et al.	375/148
2004/0120424	A1 *	6/2004	Roberts	375/327
2004/0143615	A1 *	7/2004	Yomo et al.	708/300
2005/0163238	A1 *	7/2005	Fujii	375/260
2006/0062196	A1 *	3/2006	Cai et al.	370/345
2007/0165488	A1 *	7/2007	Wildey	367/101
2009/0034589	A1 *	2/2009	Hahm et al.	375/150
2010/0189163	A1 *	7/2010	Burgi et al.	375/142
2011/0002366	A1 *	1/2011	Michaels et al.	375/148
2011/0122032	A1 *	5/2011	Sakata et al.	343/703
2011/0200126	A1	8/2011	Bontu et al.	
2013/0230312	A1 *	9/2013	Randel et al.	398/25
2013/0315342	A1 *	11/2013	Um et al.	375/295

OTHER PUBLICATIONS

Colavolpe, G., et al., Noncoherent Sequence Detection in Frequency Nonselective Slowly Fading Channels, IEEE Journal on Selected Areas in Communications, vol. 18, No. 11, Nov. 2000 pp. 2302-2311.

Wong. S.Y., et al., Multihop Localization with Density and Path Length Awareness in Non-Uniform Wireless Sensor Networks, 5 pages, IEEE 2005.

Wu, M., et al., Sequence Detection on Fading Channels without Explicit Channel Estimation, Wireless VITAE '09, pp. 370-374.

* cited by examiner

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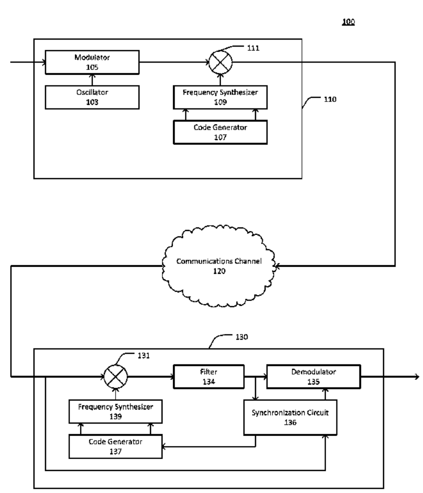
(57) **ABSTRACT**

During transmission of a signal, a decoder receives a data symbol sequence and one or more copies of the data symbol sequence. The data symbol sequence and each of the copies have unknown noise, including unknown phase shifts. The unknown phase shifts are estimated such that the data symbol sequence that was transmitted is substantially recovered.

8 Claims, 4 Drawing Sheets



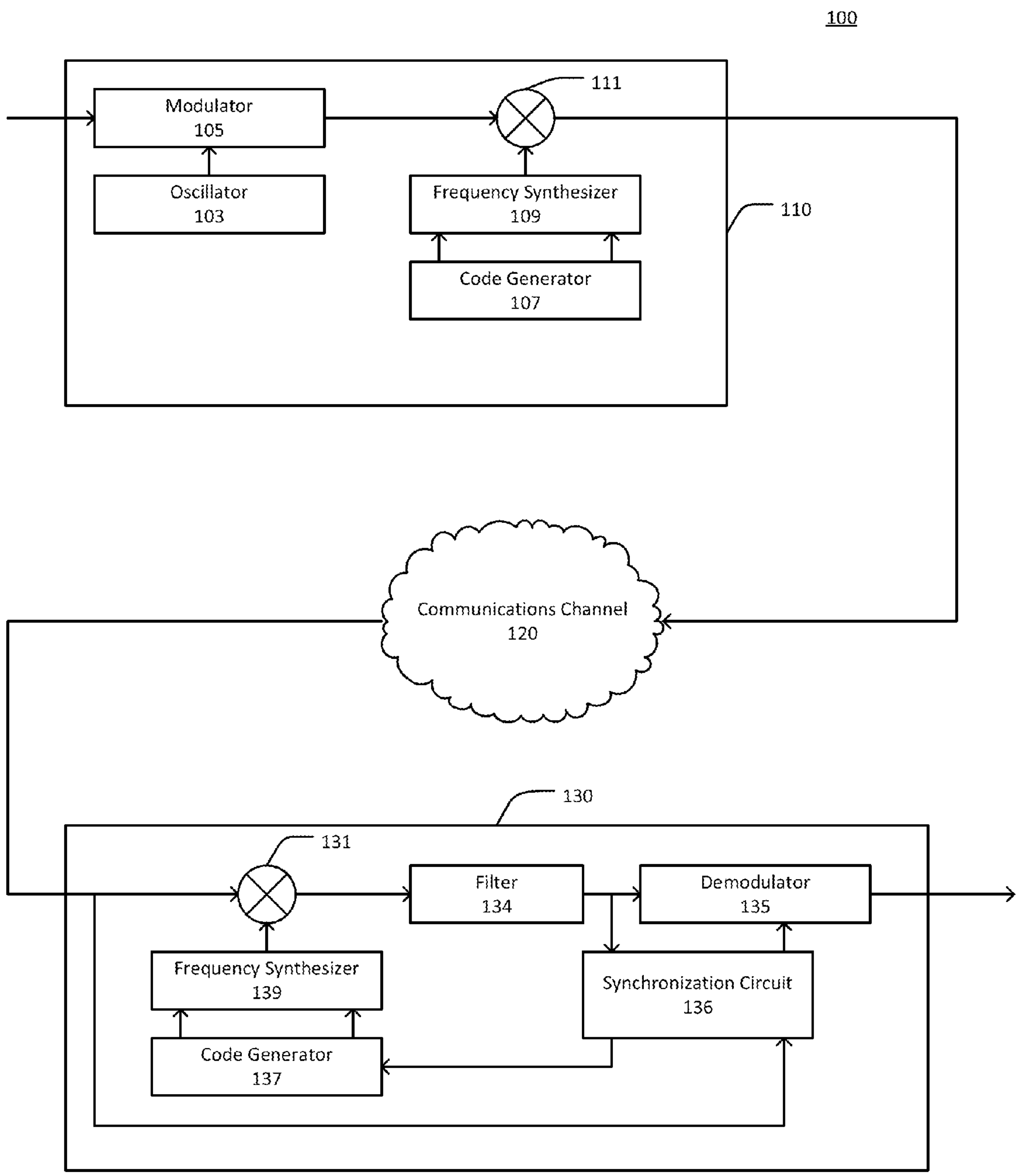

100
Modulator
105
111
Oscillator
103
Frequency Synthesizer
109
110
Code Generator
107
Communications Channel
120
130
131
Filter
134
Demodulator
135
Frequency Synthesizer
139
Synchronization Circuit
136
Code Generator
137

FIG. 1

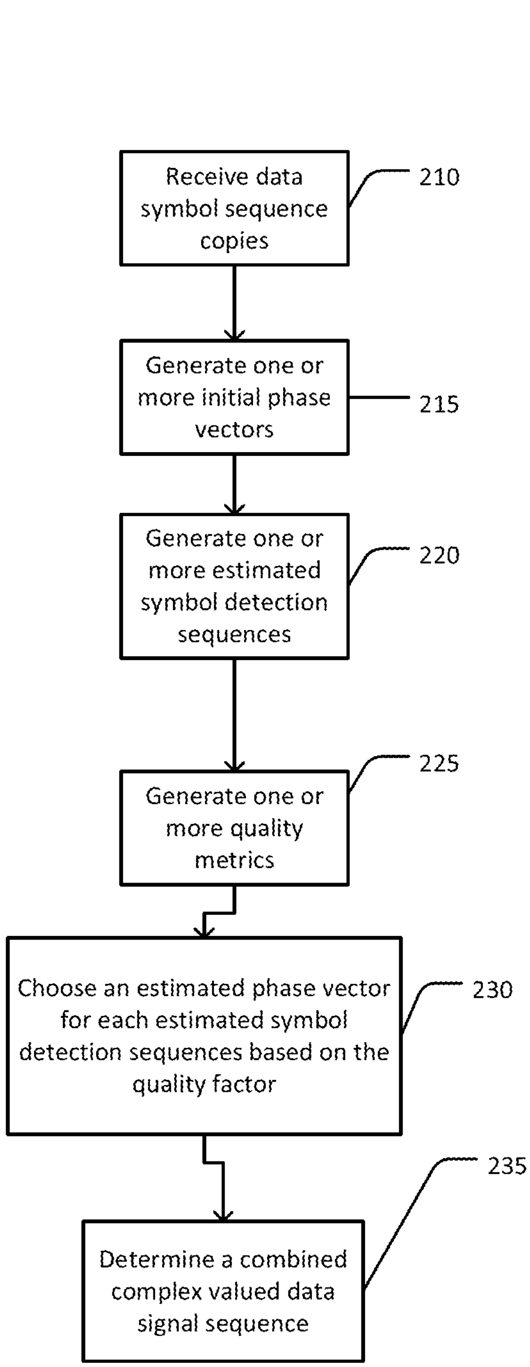

200
Receive data symbol sequence copies
210
Generate one or more initial phase vectors
215
Generate one or more estimated symbol detection sequences
220
Generate one or more quality metrics
225
Choose an estimated phase vector for each estimated symbol detection sequences based on the quality factor
230
Determine a combined complex valued data signal sequence
235

FIG. 2

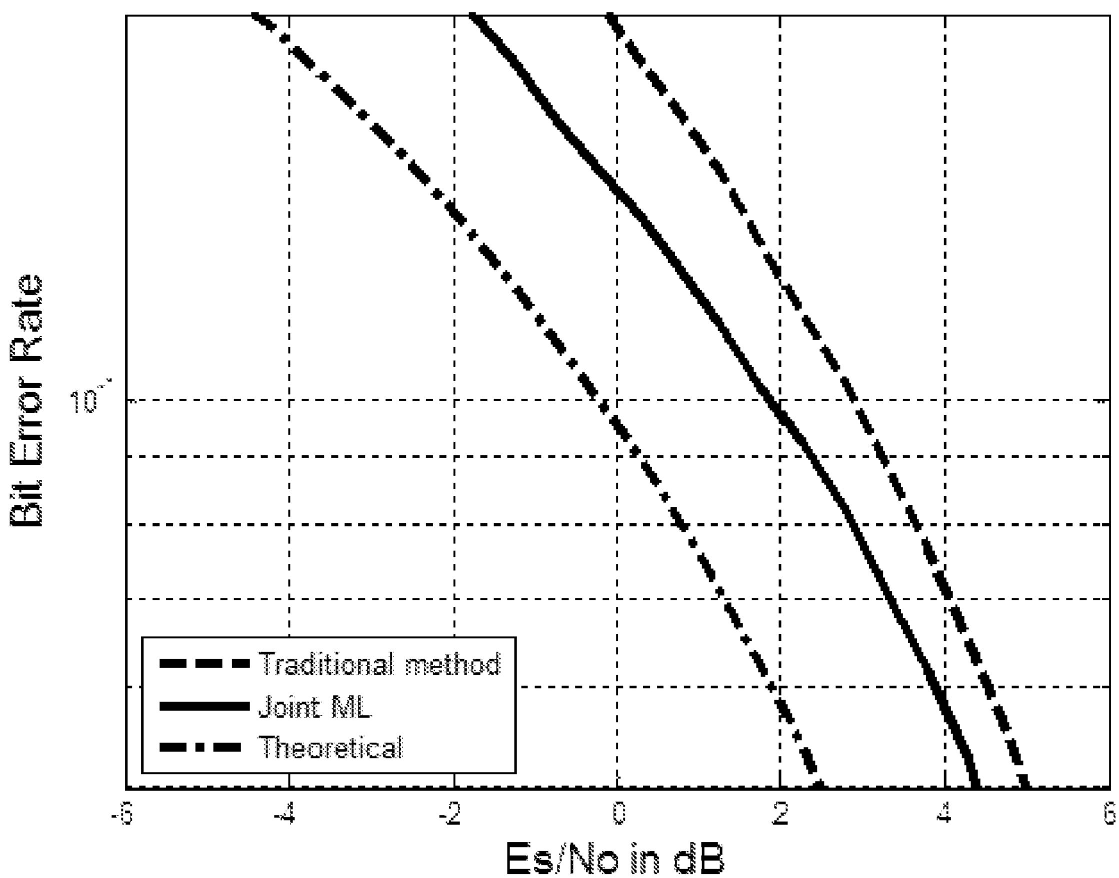

Bit Error Rate
Es/No in dB
Traditional method
Joint ML
Theoretical
-6
-4
-2
0
2
4
6

FIG 3

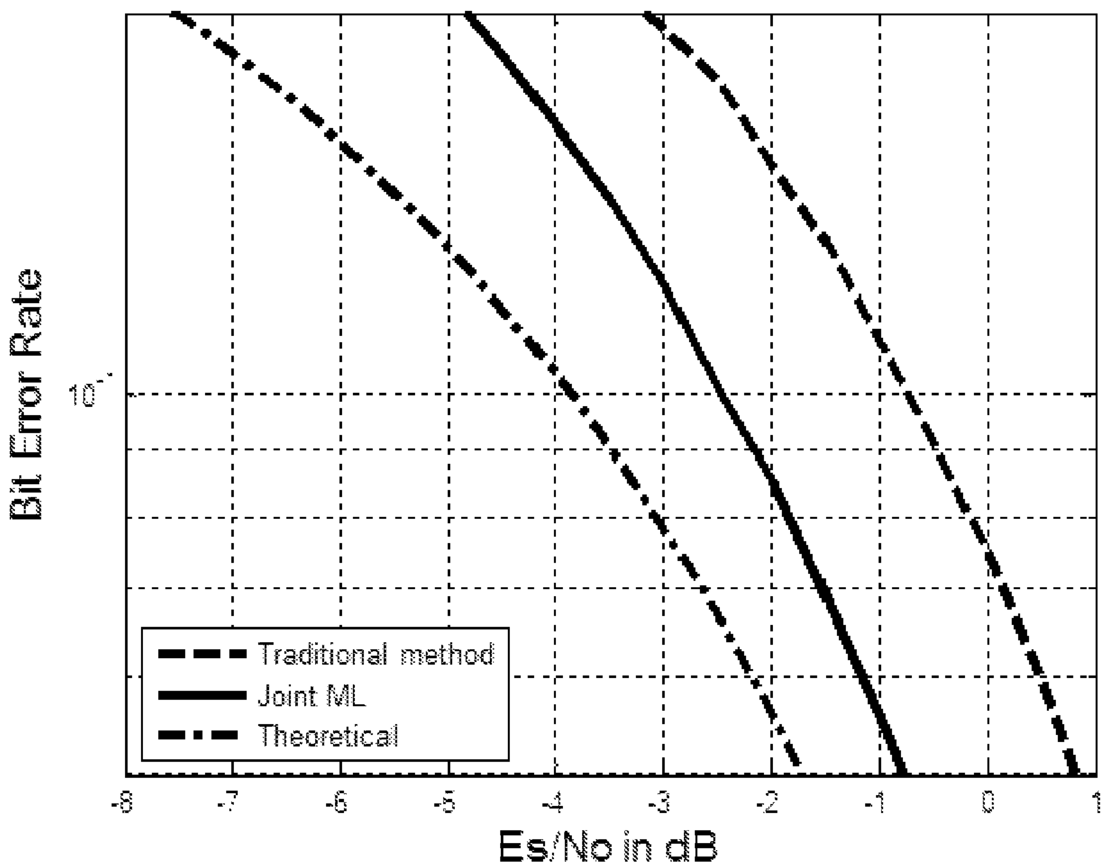

Bit Error Rate
10^-1
Traditional method
Joint ML
Theoretical
-8
-7
-6
-5
-4
-3
-2
-1
0
1
Es/No in dB

FIG 4

RECEPTION OF PHASE-SHIFTED SIGNAL SEQUENCES IN DIVERSITY RECEPTION

TECHNICAL FIELD

The invention relates generally to wireless communications. Specifically, the invention relates to transmission and reception of signals received over degraded communication channels.

BACKGROUND

Digital wireless communication can be described as a class of methods used to transmit data (or message) in the form of numbers and/or symbols from a source point and receive the data at a remote destination point. Many satellite, airborne, and terrestrial wireless systems use a digital form of communication. The digital form of communication can require that the data be transformed into symbols. The set of symbols typically used is usually small, e.g., consisting of two, four, or eight different symbols.

A larger volume of data can be represented as an ordered group (i.e., a sequence) of symbols. Each data symbol can be represented as one or more physical quantities at different points in a circuit chain of a transmitter, through a medium of air and/or free space, and in a circuit chain of a receiver, until the data is converted back to a data symbol. Some of the physical quantities can be functions over a time period.

In particular, through the medium, a data symbol can take the form of an electromagnetic wave lasting over a symbol with a particular frequency, strength, and a starting phase angle. The medium and the electromagnetic spectral band used can be together known as the communication channel. Throughout the communication system, many of the intermediate quantities corresponding to a symbol can be called signals to distinguish them from the data symbol.

These signals can experience deterministic and nondeterministic transformations as they advance from one point to the next. When the signal arrives at the receiver circuit chain, it is usually corrupted by noise and/or other forms of interference. Such interferences are examples of nondeterministic transformations that communication signals experience. The extent of interference can also be known as the level of degradation of the channel. Interference can be intentional or unintentional. Jamming a communication system by directing electromagnetic energy at the receiving antenna is an example of interference intended to disrupt communication. A channel affected by intentional interference is also known as a contested channel.

Wireless communication typically uses forward error correction (FEC) encoding to combat interference and improve the reliability (or probability) of recovering (or reconstructing) the transmitted message. FEC encoding can introduce redundant symbols (that are functions of the original data symbols) at the transmitter and typically uses a corresponding decoding scheme at the receiver. The ratio of the number of original data symbols to the total number of original plus redundant symbols is known as the coding rate. Half and one third rate FEC are common. An additional technique to improve the recovery of the original message to be transmitted over a degraded channel is to transmit multiple copies of symbols (e.g., copies of FEC encoded symbols) interspersed over a time period. A simple reasoning of how this can improve the probability of recovering the symbol is that at least one of the copies may escape interference or may be only mildly affected by interference. Such a communication method is said to use "repeat codes." Repeating FEC encoded

symbols can reduce the overall coding rate. At the receiver, signals corresponding to multiple copies of a symbol should be properly combined. In an example, let the FEC encoded data symbols needed to be transmitted be:

1, 4, 2, 2, 3, 1

At some points in the communication system, signals corresponding to a symbol can be complex numbers. In this example, let the signals corresponding to the above six symbols be the following complex numbers:

1, −i, i, i, −1, 1

at the transmitter. In the above, i denotes the unit imaginary part

$$\sqrt{-1}$$

of complex numbers. This transformation from the symbols to the signals is an example of the quadrature phase shift keying (QPSK) modulation scheme.

Let the transmitter transmit the above symbol sequence two different times in the "repeat codes" mode. The signals can experience deterministic and nondeterministic transformations as they advance through the communication system. Let the original sequence of signals be multiplied by the complex number 0.98+0.19i. The resulting signals after this transformation are:

0.98+0.19i, 0.19−0.98i, −0.19+0.98i, −0.19+0.98i, −0.98−0.19i, 0.98+0.19i

at some point in the receiver. Let the copy of the signals experience a slightly different transformation and be multiplied by the complex number 0.92+0.39i instead of being multiplied by the earlier factor 0.98+0.19i. The resulting signal sequence of the repeated transmission after this transformation is:

0.92+0.39i, 0.39−0.92i, −0.39+0.92i, −0.39+0.92i, −0.92−0.38i, 0.92+0.39i

In addition to such transformations, let the signals at the receiver be corrupted by noise over and above the transformations. Let the resulting transformed and noisy signals for the original and the repeat transmissions be

0.80+0.16i, 0.24−1.04i, −0.38+0.92i, −0.34+1.11i, −0.79−0.24i, 1.10+0.33i

and

0.87+0.31i, 0.53−0.89i, −0.43+1.05i, −0.24+0.86i, −0.83−0.45i, 0.77+0.32i

respectively. In this example, combining the two received signals sequences can be accomplished by averaging the two complex signal sequences. The resulting combined signal sequence is:

0.84+0.24i, 0.38−0.97i, −0.41+0.99i, −0.29+0.99i−0.81−0.35i, 0.94+0.33i

rounded off to two decimal places.

In the above example, the multiplication of the signal sequence corresponding to the original transmission by the complex number 0.98+0.19i and the multiplication of the signal corresponding to the repeat transmission by the complex number 0.92+0.39i can be due to the phase shifts that the signal sequences experience.

The phase shifts are examples of the above mentioned deterministic transformations in the sense that all the signals in a sequence are multiplied by the same complex number and the receiver can "anticipate" that such a multiplication takes place. It is a nondeterministic transformation in the sense that the receiver may not know the exact complex number that the signal sequence gets multiplied by. In some communication system, the circuits in the receiver can track these phase shifts fairly accurately and make the needed correction. Examples of conditions that can facilitate accurate phase tracking are (1) constant carrier frequency, (2) strong direct communication path and negligible multipath reflections of the electromagnetic waves reaching the receiver, (3) good atmospheric conditions contributing to the stability of the carrier frequency phase angle, and/or (4) adequate signal to noise ratio (SNR). Under these conditions, a system can transmit reference symbols of known values only occasionally to help the receiver with accurate phase tracking. The receiver knows the locations and values of reference symbols and uses the corresponding signals for phase tracking.

Examples of systems that do not operate with such favorable features are (1) frequency hopped systems that change the carrier frequency frequently to avoid being intercepted and also to reduce the level of channel degradation caused by narrow band jammers, and/or (2) systems operating with possible multipath receptions, with multiple antennas, and over a degraded channel. Many such systems are designed to transmit reference symbols frequently. One problem is that the reference symbols are also corrupted by noise and an estimate of the phase shift computed with the use of reference signals only can be inaccurate. Data signals carry some information about the phase shift and the receiver can use them in conjunction with the reference signals to improve the accuracy of phase shift estimation.

The problem of estimation of phase shift is compounded if the receiver can combine multiple copies of an original transmission. Systems that incorporate systematic repeat codes are not the only ones in which the receiver can take advantage of combining. There are communication systems that operate with a data link protocol in which a receiver requests a repeat if an originally received signal sequence could not be successfully converted to data symbols. Ordinarily, the receiver discards the original reception and processes only the repeated version. A proper combining of the multiple copies can be a better option. Systems that use multiple antennas can also take advantage of combining the multiple copies received from multiple antennas arising out of a single transmission.

The process of determination of symbol decisions from the received signals can be called data symbol detection (or just "detection"). Phase shift estimation and detection can be treated as a joint problem to improve the probability of correct detection and it is desirable to do so.

SUMMARY

In one aspect, the invention involves a method of detecting a data symbol sequence transmitted over a wireless communications channel. The method involves receiving, by a detector in a receiver, a plurality of data signal sequence copies. Each data signal sequence copy includes a plurality of data symbols and data-symbol phase angles, a reference signal including at least one reference symbol, and noise, and includes an unknown phase shift, wherein the data signal is approximately identical for each copy and the unknown phase shift unknown for each copy. The method also involves generating, by the detector, an initial phase vector for each data symbol, each initial phase vector is based on the plurality of data signals for one data symbol present in the plurality of signal sequence copies. The method also involves generating, by the detector, an estimated symbol detection sequence for each initial phase vector. The method also involves generating, by the detector, a quality value for each estimated symbol detection sequence. The method also involves selecting, by the detector, a global estimated phase vector, the global estimated phase vector is the estimated phase vector for the estimated symbol detection sequence having a quality value that is the highest of all quality values. The method also involves detecting, by the detector, the data symbol sequence by a) phase shifting each of the plurality of data signal sequence copies based on the global estimated phase vector, and b) combining each of the phase shifted plurality of data signal sequence copies to obtain the data symbol sequence.

In some embodiments, the method also involves iteratively updating each estimated symbol detection sequence and each estimated phase vector until each estimated phase vector is identical to a corresponding estimated phase vector of an earlier iteration.

In some embodiments, the data symbol sequence comprises a repeat code from one transmitter. In some embodiments, the data symbol sequence comprises a one-hop portion of a signal transmitted over a frequency-hopping communications channel. In some embodiments, the detector comprises a field-programmable gate array. In some embodiments, the data symbol sequence comprises a repeat code from multiple transmitters. In some embodiments, the bit error rate of the global estimated symbol detection sequence is below 10^{-5}.

In some embodiments, the method involves decoding, by a decoder receiving the combined symbol sequence from the detector, the combined symbol sequence using hard decision decoding.

In some embodiments, the method involves decoding by a decoder that can get soft decision inputs, the soft decisions being the combined signal sequence from the detector.

Advantages of exemplary embodiments include enabling a detector or detection circuit in a receiver to more efficiently combine received phase-shifted signals. Using approximately identical data signals included in multiple copies of a received symbol sequence signal, the detector can efficiently determine unknown phase errors added to each of transmitted copies when traversing the communications channel. Additional advantages include efficient combination of received copies of the transmitted signal to decrease the bit error rate of the communications system's transmission. Additional advantages include enabling the communications systems to more readily use non-coherent transmission schemes and conduct frequency hopping techniques with shorter dwell times for each hop.

BRIEF DESCRIPTION OF THE DRAWINGS

In order to better understand various exemplary embodiments, reference is made to the accompanying drawings wherein:

FIG. **1** illustrates an exemplary communications system including a receiver that receives one or more copies of a signal sequence, according to an illustrative embodiment of the technology.

FIG. **2** illustrates an exemplary method for detecting a symbol sequence from received copies of a signal sequence according to an illustrative embodiment of the technology.

FIG. **3** is a graph showing exemplary results for bit error rates (BER) versus the signal-to-noise ratios (SNR) for a receiver, according to an illustrative embodiment of the technology.

FIG. **4** is a graph showing exemplary results for bit error rates (BER) versus the signal-to-noise ratios (SNR) for a receiver, according to an illustrative embodiment of the technology.

DETAILED DESCRIPTION

FIG. **1** illustrates an exemplary communications system **100** that includes a receiver **130** that receives one or more copies of a data symbol sequence signal (e.g., a code repetition signal). The communications system **100** includes a transmitter **110**, a receiver **130**, and a communications channel **120**.

Transmitter **110** includes an oscillator **103**, a modulator **105**, a code generator **107**, a frequency synthesizer **109**, and a mixer/multiplier **111**. Each of these components of the transmitter **110** can be hardware, such as individual processors, semiconductors, field-programmable gate arrays (FPGAs), application-specific integrated circuits (ASICs) and/or other circuits. In various embodiments, the transmitter **110** is a mobile device or a fixed communications device (e.g., an antenna, phone, computer, satellite, etc.). As is apparent of one of ordinary skill in the art, transmitter **110** is exemplary and can be substituted with any suitable transmitter known in the art. The transmitter **110** receives an input signal (e.g., an analog or digital signal) and outputs a signal sequence representing a symbol sequence.

The modulator **105** receives the input signal. The modulator **105** is in communication with the oscillator **103**. The modulator **105** receives a base frequency from the oscillator **103**. The modulator **105** produces a combined signal based on the input signal and the combined signal having the base frequency. The modulator **105** can modulate the signal through various methods known in the art, such as, for example, phase-shift keying (PSK), frequency-shift keying (FSK), amplitude-shift keying (ASK), minimum-shifting keying (MSK), and/or other forms of continuous-phase modulation (CPM). In one embodiment, the modulator **105** modulates the input signal based on quadrature phase-shift keying (QPSK). In such instances, the combined signal has a data rate that is higher than the data rate as the input signal within the same bandwidth as the input signal, as the combined signal is concentrated around four phases uses by the modulator **105**.

The modulator **105** outputs the combined signal. The modulator is in communication with the mixer **111**. The mixer **111** takes as input the combined signal. The mixer **111** is also in communication with the frequency synthesizer **109**. The mixer **111** determines the symbol sequence signal based on the input signal and the output of the frequency synthesizer **109**. The mixer **111** outputs the symbol sequence signal to the communications channel **120**.

The frequency synthesizer **109** is in communication with mixer **111** and code generator **107**. The frequency synthesizer outputs a carrier frequency for a hopset to the mixer **111**. The carrier frequency for the hopset is used by mixer **111** to determine a frequency-hopping pattern for the modulate signal output from the modulator **105**. The carrier frequency in the hopset is determined based on a code segment output by the code generator **107**.

The frequency synthesizer **109** receives the code segment from the code generator **107**. The code generator **107** determines a pseudorandom code to supply as input to the frequency synthesizer **109**. The pseudorandom code defines the hopset. In some embodiments, the code generator **107** can receive as input a secret key and a number from a number generator (such as a time-of-day clock) to generate the pseudorandom code. In such instances, the frequency synthesizers **109** and code generator **107** of the transmitter **110** can be identical to a transmit frequency synthesizer **139** and a code generators **137** of the receiver **130** so that the transmitter **110** and receiver **130** use an identical key (e.g., a shared private key).

In operation, the transmitter **110** receives the input signal. The transmitter **110** codes the input signal into a symbol sequence signal. In some embodiments, the transmitter **110** can include an encoder (not shown) that produces the symbol sequence signal based on the input signal and/or predefined parameters. In some embodiments, the transmitter **110** receives the input signal is the symbol sequence signal. The transmitter **110** transmits the symbol sequence signal through communications channel **120** to receiver **130**.

In some embodiments, the transmitter **110** transmits the symbol sequence signal as a spread-spectrum signal having a frequency hopping pattern within a specified hopband. For example, the communication system can function with each sub-band of 25 KHz hopping over numerous such sub-bands in the wide band of 30-60 MHz.

In some embodiments, the communications system **100** has multiple transmitters **110** that send signals to receiver **130**, such as in a multiple-input, single-output (MISO) system.

In some embodiments, the transmitter **110** employs modulation and frequency-hopping techniques in concert to send to the receiver **130**. For example, the transmitter **110** can employ phase-shift keying (e.g., QPSK or 8-PSK) to alter the phase of the input signal, while at the same time, set a carrier frequency in the hopset, such that transmitter **110** can send copies of the spread-spectrum signal sent at the chosen carrier frequency within the hopset at different phases for a specified time (dwell time) before hopping to another carrier frequency in the hopset based on the chosen frequency-hopping pattern.

The communications channel **120** is in communication with the transmitter **110** and the receiver **130**. The communications channel **120** can be any wired or wireless communication mechanism know in the art. For example, the communications channel **120** can be a "WiFi" communication channel having a frequency range, such as 2412-2484 MHz.

The communications channel **120** can add noise to the signals transmitted from the transmitter **110**. Communication channel **120** can also receive noise, such as additive white Gaussian noise (AWGN) as an additional input (not shown) to the signals sent from transmitter **110**. The additive noise in the communications channel **120** can degrade the signal. In some embodiments, the noise added to the signal in communications channel **120** is random and independent, changing the amplitude and phase of the originally-sent signal. In some embodiments, third parties can use devices such as interceptors or frequency jammers to interfere with communications between transmitter **110** and receiver **130**.

The receiver **130** is in communication with the communications channel **120**. The receiver includes a mixer/divider **131**, a filter **134**, a demodulator **135**, a synchronization circuit **136**, a code generator **137**, and a frequency synthesizer **139**. The components of receiver **130** can be hardware such as individual processors, semiconductors, field-programmable gate arrays (FPGAs), application-specific integrated circuits (ASICs) and/or other circuits. The receiver **130** can be a mobile or fixed communications device (e.g., an antenna, phone, computer, satellite, etc.). The receiver **130** can be any receiver known in the art.

The mixer/divider **131** is in communication with communications channel **120**, frequency synthesizer **130** and filter **134**. The mixer/divider **131** receives as input as signal from the communications channel **120** and a frequency from the frequency synthesizer **130**.

The frequency synthesizer **130** is in communication a code generator **137** and a synchronization module **136**. The frequency synthesizer **130** outputs a carrier frequency for a hopset to the mixer **131**. The carrier frequency for the hopset can be based on output from the synchronization module **136**, output from the code generator **137**, and/or identical to the carrier frequency for the hopset determined by the frequency synthesizer **109** in the transmitter **110**.

The frequency synthesizer **139** receives the code segment from the code generator **137**. The code generator **137** can determine a pseudorandom code to supply as input to the frequency synthesizer **139**. The pseudorandom code defines the hopset. In some embodiments, the code generator **137** can receive as input a secret key and a number from a number generator (such as a time-of-day clock) to generate the pseudorandom code. In some embodiments, the code generator **137** generates a code identical to the code generated by the code generator **107** of the receiver. In some embodiments, the transmitter **110** and receiver **130** use an identical key (e.g., a shared private key).

The filter **134** is in communication with the mixer divider **131** and the synchronization circuit **136**. The mixer/divider **131** removes the frequency hopping pattern (using a frequency from frequency generator **139** as an input). The filter **134** receives the de-mixed signal and can remove extraneous elements from the signal, such as double frequency components and/or power outside the intended carrier frequency. In some embodiments, filter **134** is a bandpass filter that produces a dehopped signal for demodulator **135** and synchronization circuit **136**.

The demodulator **135** is in communication with the filter **134** and synchronization circuit **136**. The demodulator **135** extracts the symbol sequence signal from the dehopped signal received from the filter **133**. The demodulator **135** can employ techniques for demodulation that are capable of demodulating the modulated signals. For example, if transmitter **110** includes a modulator **105** that uses QPSK for modulating the spread-spectrum signal it transmits, receiver **130** includes a demodulator **135** that uses QPSK to demodulate received copies of the spread-spectrum signal.

The synchronization module **136** is in communication with the communications channel **120**, the demodulator **135**, filter **134**, and the code generator **137**. The synchronization module **136** synchronizes the frequency hopping pattern of received copies of the symbol sequence signal. For example, the synchronization circuit **136** can receive a dehopped signal from filter **134** and the spread-spectrum signal from the communications channel **120** as input and output control signals to the demodulator **135** and the code generator **137**.

In some embodiments, the receiver **130** can receive multiple copies of the same symbol sequence signal segment and combine the copies to efficiently determine the data component of the segment. In such instances, combining multiple copies of the same segment can eliminate the effects of noise, fading, and intentional interference experienced by the symbol sequence signal segment when traveling through communication channel **120**.

In some embodiments, the receiver **130** can receive the symbol sequence signal as a spread-spectrum signal at a specific carrier frequency within a frequency-hopping pattern used by transmitter **110**. In some embodiments, the receiver **130** receives low-rate codes as known in the art.

In some embodiments, the receiver **130** can also include a decoder (not shown). The decoder can receive the recovered symbol sequence output by demodulator **135** to determine the information stored in the symbol sequence. For example, receiver **110** can include a decoder to decode the symbol sequence signal employing decoding techniques to decode signals encoded based on, such as, Reed-Solomon (RS) codes, Forward Error Correction (FEC) codes, Viterbi codes, turbo codes, and Low-Density Parity Codes (LPDC). In some embodiments, the decoder receives the symbol sequence as specific values (i.e., hard-decision detection), while in other embodiments, the decoder receives the symbol sequence as Log Likelihood Ratios (LLRs) for individual bits in the symbol sequence (i.e., soft-decision detection).

FIG. **2** illustrates an exemplary method **200** for detecting a symbol sequence signal from a plurality of received copies of a symbol sequence transmitted over a wireless communications channel, according to an illustrative embodiment of the technology.

The method involves receiving a data symbol sequence and one or more of copies of the symbol sequence (Step **210**) (e.g., receiving by receiver **130**, as described above in FIG. **1**). The received data symbol sequence and the copies of the data symbol sequence can be referred to generally as data symbol sequence copies. The data symbol sequence copies each include a data signal, a reference signal, and noise. The data signal can be one or more data symbols and one or more data-symbol phase angles. The data signal can be a complex number. For example, a data symbol sequence having 100 data signals can be transmitted four times. In this example, there are four data symbol sequences (one original and three copies) and 400 data signals in total.

Each copy of the data symbol sequence is substantially identical to the original data symbol sequence, except for the noise and except for an unknown phase shift. For example, the receiver can receive a data symbol sequence of a spread-spectrum signal and three copies, each of the data symbol sequence and the three copies can have independent phase shifts and AWGN.

Detecting a symbol sequence signal can include determining a globally optimal phase shift vector ϕ and a corresponding symbol detection sequence $\{\theta\}$. The globally optimal phase shift vector ϕ and the corresponding symbol detection sequence $\{\theta\}$ can be determined by maximizing:

$$\max_{\varphi,\{\theta\}} f(a, b, s, t, \varphi, \theta) = \max_{\varphi,\{\theta\}} \sum_{j=1}^{c}\left[\left\{\sum_{i=1}^{n} a_{ij}\cos(\varphi_j - b_{ij} + \theta_i)\right\} + \left\{\sum_{i=1}^{r} s_{ij}\cos(\varphi_j - t_{ij})\right\}\right]. \quad \text{EQN. 1}$$

where c is the number of copies, n is the number of data symbols in the symbol sequence, r is the number of reference symbols for a total of n+r symbols in the sequence of symbols transmitted a_{ij} is the magnitude of the i-th data symbol in the j-th received copy, b_{ij} is the phase angle of the i-th data symbol in the j-th received copy, ϕ_j is the unknown phase shift of the j-th copy, s_{ij} is the magnitude of the i-th reference symbol in the j-th copy and t_{ij} is the phase angle of the same. In the maximization, all possible decisions can be considered for each θ_i. For example, for the QPSK modulation scheme, the possible values for each θ_i can be 0,

$$\frac{\pi}{2},$$

π,

$$\frac{3\pi}{2}$$

radians. In some embodiments, the magnitudes of the received copies of data signal sequences can be scaled differently prior to determining the phase shift vector. For example, the received signals can be first normalized to have the same statistical variance and the combining factors ρ_j can be evaluated for maximal ratio combining as described in Linear Diversity Combining Techniques by D. G. Brennan, Proceedings of the IRE, June 1959, which in incorporated herein in its entirety. If x_{ij}, j=1, . . . , c and i=1, . . . , n are the magnitudes of the received data signals and y_{ij}, j=1, . . . , c and i=1, . . . , r are the magnitudes of the received reference signals, the normalized signal magnitudes to be used in EQN. 1 can be obtained as

$$a_{ij}=\rho_j x_{ij}, j=1, \ldots, c \text{ and } i=1, \ldots, n$$

and

$$s_{ij}=\rho_j y_{ij}, j=1, \ldots, c \text{ and } i=1, \ldots, r$$

Determining the globally optimal phase shift vector ϕ by maximizing EQN. **1** shown above can involve a computationally intensive number of options to consider. In some embodiments, the globally optimal phase shift vector ϕ can be determined by approximating the maximization of EQN. **1** above.

Approximating the maximization of EQN. **1** can involve generating one initial phase vector for each data symbol (Step **215**). For example, for a data symbol sequence having one hundred data symbols transmitted four times (e.g., one original and three copies), one hundred initial phase vectors are generated.

In some embodiments, each initial phase vector can be determined by selecting an optimum phase vector from a list of potential initial phase vectors. Each initial phase vector in the list of potential initial phase vectors can be determined by maximizing:

$$h_i = \max_{\varphi_{ij},\theta_k} \sum_{j=1}^{c} \left\{ a_{ij}\cos(\varphi_{ij} - b_{ij} + \theta_k) + \sum_{l=1}^{r} s_{lj}\cos(\varphi_{ij} - t_{lj}) \right\} \quad \text{EQN. 2}$$

EQN. 2 can be expressed as a linear combination of $\cos(\phi_{ij})$ and $\sin(\phi_{ij})$ as follows:

$$h_{ik} = \sum_{j=1}^{c} \left\{ a_{ij}\cos(\varphi_{ijk} - b_{ij} + \theta_k) + \sum_{l=1}^{r} s_{lj}\cos(\varphi_{ijk} - t_{lj}) \right\} \quad \text{EQN. 3}$$

$$= \sum_{j=1}^{c} \cos(\varphi_{ijk}) \left(a_{ij}\cos(b_{ij} - \theta_k) + \sum_{l=1}^{r} s_{lj}\cos(t_{lj}) \right) + \sin(\varphi_{ijk}) \left(a_{ij}\sin(b_{ij} - \theta_k) + \sum_{l=1}^{r} s_{lj}\sin(t_{lj}) \right) \quad \text{EQN. 4}$$

In some embodiments, an intermediate parameter ψ_k is determined as follows:

$$\cos(\psi_{ijk}) = \frac{A_{ijk}}{C_{ijk}}, \text{ and} \quad \text{EQN. 5}$$

$$\sin(\psi_{ijk}) = \frac{B_{ijk}}{C_{ijk}}. \quad (20)$$

where A_{ijk} are intermediate parameters, B_{ijk} are intermediate parameters, and C_{ijk} are intermediate parameters. In some embodiments, the intermediate parameters A_{ijk} are determined as follows:

$$A_{ijk} = \left(a_{ij}\cos(b_{ij} - \theta_k) + \sum_{l=1}^{r} s_{lj}\cos(t_{lj}) \right), \quad \text{EQN. 6}$$

In some embodiments, the intermediate parameters B_{ijk} are determined as follows:

$$B_{ijk} = \left(a_{ij}\sin(b_{ij} - \theta_k) + \sum_{l=1}^{r} s_{lj}\sin(t_{lj}) \right), \quad \text{EQN. 7}$$

In some embodiments, the intermediate parameters C_{ijk} are determined as follows:

$$C_{ijk} = \sqrt{A_{ijk}^2 + B_{ijk}^2}, \quad \text{EQN. 8}$$

In some embodiments, applying EQNs 5-8 to EQN. 3 results in:

$$h_{ik} = \sum_{j=1}^{c} C_{ijk}\cos(\varphi_{ijk} - \psi_{ijk}), \quad \text{EQN. 9}$$

$$h_i = \max_{\varphi_{ij},k} \sum_{j=1}^{c} C_{ijk}\cos(\varphi_{ijk} - \psi_{ijk}). \quad \text{EQN. 10}$$

In some embodiments, the j-th component of the initial phase vector corresponding to data symbol i can be determined as follows:

$$\phi_{ij}=\psi ijk^*. \quad \text{EQN. 11}$$

where k* is the value of k that maximizes the argument of EQN. 10. For example, for k between 1 and 10, k* is the k value between 1 and 8 that maximizes EQN. 10. In some embodiments, each initial phase shift vector is determined by:

$$\Phi_i=(\phi_{i1}, \ldots, \phi_{ic}), i=1, \ldots, n. \quad \text{EQN. 12}$$

where c is the number of copies of the signal sequence and i is the position of the data signal for which the initial phase shift vector Φ_i is evaluated in EQN. 12

The method also involves determining an estimated symbol detection sequence for each initial phase vector (Step **220**). For example, for signal sequence copies, each copy having one data signal, one hundred initial phase shift vectors are determined. For each initial phase shift vector, a corresponding estimated symbol detection sequence can be determined, thus, in this example, one hundred estimated symbol detection sequences are determined. In some embodiments, each estimated symbol detection sequence is determined by:

$$f_k(\Phi_i, \chi_{ik}) = \max_{\theta_k} \sum_{j=1}^{c} a_{kj}\cos(\varphi_{ij} - b_{kj} + \theta_k). \quad \text{EQN. 13}$$

where k is the position of the data signal for which the detection is being estimated using the initial phase vector Φ_i and the maximization is carried out over all possible values of θ_k and χ_{ik} is the value of θ_k that maximizes the right hand side of EQN. 13. χ_{ik} can be the estimated symbol decision (or detection) for the data symbol position k using the initial phase vector Φ_i.

The method also involves generating a quality metric for each symbol sequence detection (Step **225**). In some embodiments, the quality metric for each symbol sequence is determined by maximizing EQN. 14:

$$f(\varphi)=\sum_{j=1}^{c}\left\{\cos(\varphi_j)\left[\sum_{i=1}^{n}a_{ij}\cos(b_{ij}-\theta_i)+\sum_{i=1}^{r}s_{ij}\cos(t_{ij})\right]+\sin(\varphi_j)\left[\sum_{i=1}^{n}a_{ij}\sin(b_{ij}-\theta_i)+\sum_{i=1}^{r}s_{ij}\sin(t_{ij})\right]\right\}. \quad \text{EQN. 14}$$

with respect to the variable phase shift vector ϕ where a_{ij}, b_{ij}, s_{ij}, t_{ij} are values of received signals as defined above, θ_i is the estimated detection of in the i-th data signal position, and ϕ_j is the j-th component of ϕ.

In some embodiments, an intermediate parameter ψ_j is determined as follows:

$$\sin(\psi_j)=\frac{B_j}{C_j},\ \text{and} \quad \text{EQN. 15}$$

$$\cos(\psi_j)=\frac{A_j}{C_j}. \quad \text{EQN. 16}$$

where A_j is an intermediate parameter, B_j is an intermediate parameter and C_j is an intermediate parameter. In some embodiments, the intermediate parameter A_j is determined as follows:

$$A_j=\left[\sum_{i=1}^{n}a_{ij}\cos(b_{ij}-\theta_i)+\sum_{i=1}^{r}s_{ij}\cos(t_{ij})\right], \quad \text{EQN. 17}$$

In some embodiments, the intermediate parameter B_j is determined as follows:

$$B_j=\left[\sum_{i=1}^{n}a_{ij}\sin(b_{ij}-\theta_i)+\sum_{i=1}^{r}s_{ij}\sin(t_{ij})\right], \quad \text{EQN. 18}$$

In some embodiments, the intermediate parameter C_j is determined as follows:

$$C_j=\sqrt{A_j^2+B_j^2}, \quad \text{EQN. 19}$$

In some embodiments, $\Sigma_{j=1}^{c} C_j$ is the quality metric for the estimated detection sequence. In the example of one hundred data symbols and four copies, there are one hundred quality metric values.

The method also involves selecting the estimated symbol detection sequence and the phase shift vector that has the highest quality factor among all the quality factors (Step **230**). Selecting the estimated symbol detection sequence can involve choosing an estimated phase vector for each of the estimated symbol detection sequence. In some embodiments, the estimated phase vector for the estimated symbol detection sequence is determined by selecting:

$$\phi^*_j=\psi_j,\ j=1,\ldots,c. \quad \text{EQN. 21}$$

where ϕ^*_j is the estimated phase shift value for the j-th of the c copies. The estimated phase shift vector for the estimated symbol sequence under consideration is the vector of these values, and can be denoted as follows:

$$\phi^*=[\phi^*_1,\ldots,\phi^*_c \quad \text{EQN. 22}$$

The estimated phase shift vector along with its quality factor for each estimated symbol detection sequence is determined. In the example of one hundred data symbols and four copies of signal sequences, there are one hundred estimated detection sequences, each with its own quality factor and its own estimated phase shift vector.

The method also involves determining a combined complex valued data signal sequence (Step **235**). The combined complex valued data signal sequence can be determined as follows:

$$z_k=\sum_{j=1}^{c}a_{kj}[\cos(b_{kj}-\varphi_j)+i\sin(b_{kj}-\varphi_j)] \quad \text{EQN. 23}$$

where z_k is the combined complex signal for the k-th of the n data symbols, a_{kj} and b_{kj} are the magnitude and phase angle of the j-th copy of the k-th of the n data signals, ϕ_j is the estimated phase shift for the j-th copy with the highest quality factor, and

$$i=\sqrt{-1}$$

is the unit imaginary number.

In some embodiments, the combining is done based on a weighted sum. The weighted sum can be determined as follows:

$$z_k=\sum_{j=1}^{c}\eta_j a_{kj}[\cos(b_{kj}-\varphi_j)+i\sin(b_{kj}-\varphi_j)] \quad \text{EQN. 24}$$

where the weighting factor for the j-th copy is η_j.

FIG. **3** shows a graph **300** of bit error rates (BER) versus signal-to-noise ratios (SNR) for a detector, according to illustrative embodiments of the technology. The SNR used here is the symbol energy per unit noise power spectral density, denoted by Es/No. The BER for four copies, 8-PSK modulations, and two (2) reference symbols and twelve (12) data symbols in each copy, denoted by 2+12 is shown for typical prior art method, for the method according to an exemplary embodiments of the invention, and a coherent theoretical approach. The coherent theoretical approach can be applicable if, for example, there are no unknown phase shifts and if the receiver can take advantage of it.

FIG. **4** shows a graph **400** of bit error rates (BER) versus signal-to-noise ratios (SNR) for a detector, according to illustrative embodiments of the technology. The SNR in FIG. **4** is the symbol energy per unit noise power spectral density, denoted by Es/No. The BER for four copies, QPSK modulations, and three (3) reference symbols and fifty four (54) data symbols in each copy, denoted by 3+54 is shown for typical prior art method, for the method according to an exemplary embodiment of the invention, and a coherent theoretical approach. The coherent theoretical approach can be applicable if, for example, there are no unknown phase shifts and if the receiver can take advantage of it.

It should be apparent from the foregoing description that various exemplary embodiments of the invention may be implemented in hardware and/or firmware. Furthermore, various exemplary embodiments may be implemented as instructions stored on a machine-readable storage medium, which may be read and executed by at least one processor to perform the operations described in detail herein. A machine-readable storage medium may include any mechanism for storing information in a form readable by a machine, such as a personal or laptop computer, a server, or other computing device. Thus, a machine-readable storage medium may include read-only memory (ROM), random-access memory (RAM), magnetic disk storage media, optical storage media, flash-memory devices, and similar storage media.

It should be appreciated by those skilled in the art that any block diagrams herein represent conceptual views of illustrative circuitry embodying the principals of the invention. Similarly, it will be appreciated that any flow charts, flow diagrams, state transition diagrams, pseudo code, and the like represent various processes which may be substantially represented in machine readable media and so executed by a computer or processor, whether or not such computer or processor is explicitly shown.

Although the various exemplary embodiments have been described in detail with particular reference to certain exemplary aspects thereof, it should be understood that the invention is capable of other embodiments and its details are capable of modifications in various obvious respects. As is readily apparent to those skilled in the art, variations and modifications can be affected while remaining within the spirit and scope of the invention. Accordingly, the foregoing disclosure, description, and figures are for illustrative purposes only and do not in any way limit the invention, which is defined only by the claims.

What is claimed is:

1. A method of detecting a data symbol sequence transmitted over a wireless communications channel, the method comprising:

receiving, by a detector in a receiver, a plurality of data symbol sequence copies, each data symbol sequence copy comprising:

a data signal sequence including a plurality of data symbols and a data-symbol phase angle,

a reference signal including at least one reference symbol; and

noise and an unknown phase shift, wherein the data signal sequence is approximately identical for each copy and the unknown phase shift unknown for each copy;

generating, by the detector, an initial phase vector for each data signal in the plurality of data signal sequence copies, each initial phase vector is based on the plurality of data signal sequence copies;

generating, by the detector, an estimated symbol detection for each data signal in each of the plurality of data symbol sequence copies, each estimated symbol detection sequence is based on a corresponding initial phase vector;

generating, by the detector, a quality value for each estimated symbol detection sequence;

selecting, by the detector, a global estimated phase vector, the global estimated phase vector is the estimated phase vector for the estimated symbol detection sequence having a quality value that is the highest of all quality values; and

detecting, by the detector, the data symbol sequence by:

a) phase shifting each of the plurality of data symbol sequence copies based on the global estimated phase vector, and

b) combining each of the phase shifted plurality of data signal sequence copies to obtain the data symbol sequence.

2. The method of claim **1**, further comprising:

iteratively updating each estimated symbol detection sequence and each estimated phase vector until each estimated phase vector is identical to a corresponding estimated phase vector of an earlier iteration.

3. The method of claim **1**, wherein the data symbol sequence comprises a repeat code from

one transmitter.

4. The method of claim **3**, wherein the data symbol sequence comprises a one-hop portion

of a signal transmitted over a frequency-hopping communications channel.

5. The method of claim **1**, wherein the detector comprises a field-programmable gate array.

6. The method of claim **1**, wherein the data symbol sequence comprises a repeat code from

multiple transmitters.

7. The method of claim **1**, wherein a bit error rate of the global estimated symbol detection sequence is below lo-s.

8. The method of claim **1**, further comprising:

decoding, by a decoder receiving the combined symbol sequence from the detector, the combined symbol sequence using hard decision decoding.

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